Conversion Gain and Noise of Niobium Superconducting Hot-Electron-Mixers

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Abstract— **A study has been done of microwave mixing at 20 GHz using the nonlinear (power dependent) resistance of thin niobium strips in the resistive state. Our experiments give evidence that electron-heating is the main cause of the nonlinear phenomenon. Also a detailed phenomenological theory for the determination of conversion properties is presented. This theory is capable of predicting the frequency-conversion loss rather accurately for arbitrary bias by examining the I–V-characteristic. Knowing the electron temperature relaxation time, and using parameters derived from the I–V-characteristic also allows us to predict the –3-dB IF bandwidth. Experimental results are in excellent agreement with the theoretical predictions. The requirements on the mode of operation and on the film parameters for minimizing the conversion loss (and even achieving conversion gain) are discussed in some detail. Our measurements demonstrate an intrinsic conversion loss as low as 1 dB. The maximum IF frequency defined for –3-dB drop in conversion gain, is about 80 MHz. Noise measurements indicate a device output noise temperature of about 50 K and SSB mixer noise temperature below 250 K. This type of mixer is considered very promising for use in low-noise heterodyne receivers at THz frequencies.**

I. Introduction

For radio astronomy and remote sensing applications in the THz region there is a strong need for receivers with much higher sensitivity than is available at present. Superconductor Hot-Electron Bolometer (HEB) mixers utilizing thin superconducting films in the resistive state have recently emerged as a serious alternative to the traditional mixers used in THz receivers [1], [3]. Today, most receivers for frequencies near and above 1 THz have to rely on Schottky-diode mixers, with rather poor sensitivity [4], [5]. Low noise SIS mixers with excellent performance, have replaced Schottky-diode mixers for frequencies up to about 650 GHz, corresponding to the energy gap of niobium [4], [6]. Since niobium tri-layer technology is by far the most successful SIS-mixer technology and since the RF loss may be significant above the energy gap of niobium [7], it may be very difficult to realize low noise SIS mixers near or above about 700 GHz. InSb hot electron mixers have high sensitivity at frequencies around 1 THz, but the narrow bandwidth (≈1 MHz) limits its possible applications [8]–[10]. A more recent development where a 2-D electron gas in a heterostructure is used for hot electron mixing allows IF bandwidths of more than one GHz and may possibly be competitive at THz frequencies [11], [12]. Another interesting device is the diffusion-cooled HEB, which is reported to have a bandwidth of at least 2 GHz [13], [14].

In this paper we have examined a Superconductor HEB mixer, consisting of narrow strips of niobium-film biased in the resistive state. This recently proposed device [1], [3] has a very simple structure, and can be realized with a technology which is relatively uncomplicated compared to the SIS and Schottky diode fabrication. Our experiments at 20 GHz show very promising results, i.e., low conversion loss and noise temperature. Based on the frequency insensitive nature of the electromagnetic interaction and the non linearity of these devices, these results should be possible to translate to THz frequencies, where the device can be integrated with planar antennas in Quasioptical systems.

The purpose of the study presented in this paper has been to carefully measure and model conversion loss and noise properties of the HEB at a low millimeter wave frequency, in order to establish a firm basis for the design of future THz HEB mixers.

II. The Device

A superconducting hot electron bolometer consists of one or several superconducting thin film strips, deposited on a substrate, of for example, silicon, single crystalline quartz or sapphire. The strips are cooled to the superconducting state and then heated by dc and microwave power to the neighborhood of the superconducting to normal transition-temperature, T_c, here called the resistive state, where the superconductor will gradually become normal (Fig. 1).

The resistance of the device in the resistive state, may be explained with the help of several possible physical phenomena, such as formation of normal domains, phase slip centers, and moving magnetic vortices. Let us adopt a simplistic model:[1] assume one resistive region (or several that may be combined into one). The electron temperature, θ, in the "resistive transition regions" is in the interval ΔT_c, marked in Fig. 1.

The main power dependence of the resistance may be due to creation and annihilation of "hot spots" and heating of

Manuscript received July 11, 1994; revised November 15, 1994. This work was supported by the Swedish National Board of Industrial and Technical Development (NUTEK) and Grant NAF000 from the International Science Foundation. B. Karasik received support from the Russian Council on High-Tc Problem under Grant 93169 and from the Swedish Institute.

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IEEE Log Number 9409124.

[1] Note that the validity of the circuit-based model which we will present in the next section does not depend on the details of the microscopic model, such as the one suggested here.

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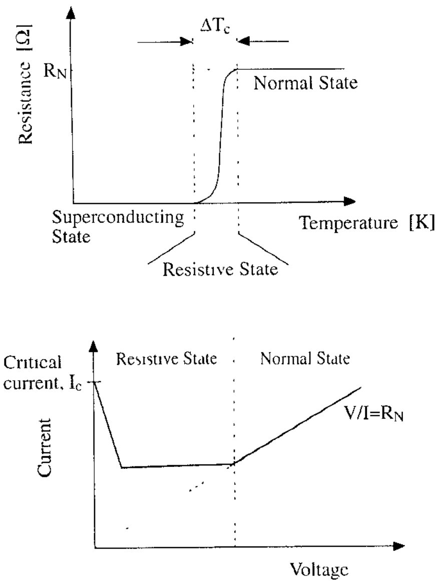


Fig. 1 The three states of the bolometer.

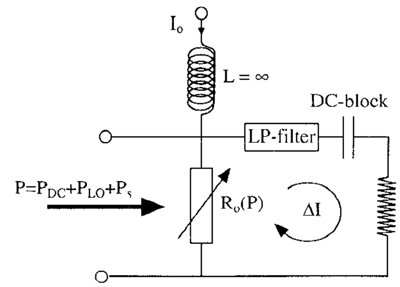


Fig. 2. Equivalent circuit of bolometer with load.

electrons in the resistive state. Here the gradient in the electron temperature is also steepest. The electron temperature as well as the lattice temperature of the superconductor outside these areas are low enough to allow zero resistance.

Note the energy gap of the superconductor, 2Δ. If the signal and LO frequency are high so that $f_{\rm LO}, f_s > 2\Delta/h$, the full length of the superconducting strip will be seen as a normal conductor strip. For $f_{\rm LO}, f_s < 2\Delta/h$ the RF-resistance will depend on the bias point and the amount of normal regions in the strip. If the intermediate frequency, $f_{\rm IF}$, is such that $f_{\rm IF} < 2\Delta/h$, it will only develop a voltage across the normal regions while the superconducting part is simply a short-circuit. These facts have important consequences on the functions of the mixer.

The maximum IF-frequency is determined by the bolometer response time to an external perturbation, i.e., the electron temperature relaxation time τ_θ, thus $f_{\rm IF} < 1/(2\pi\tau_\theta)$. If the strips are made sufficiently thin, the time constant for phonons in the superconductor to escape to the substrate, τ_{ph-s}, may become shorter than the time constant for phonon electron interaction in the superconductor, τ_{ph-e}. If the strips are made narrow, then the back flow of phonons from substrate to the superconductor will be reduced, and thus τ_{s-ph} will be long. At the same time [15], for Nb where $C_e \gg C_p$ at all temperatures below 10 K, τ_{ph-e} is much shorter than the electron phonon relaxation, τ_{e-ph} (from the detailed balance equation $C_e/\tau_{e-ph} \approx C_p/\tau_{ph-e}$, where C_e and C_p are the specific heats of the electrons and phonons respectively). Consequently it is possible to heat the electrons *above* the temperature of the lattice and τ_θ will be dominated by τ_{e-ph}, which is not the case for ordinary bolometers where the electrons and the lattice are heated to the same temperature.

In practice, to avoid lattice heating effects in the superconductor, the bolometer should be less than or about 100 Å thick and about 1 μm wide. For Nb the resulting bandwidth corresponding to $(2\pi\tau_{e-ph})^{-1}$ is of the order 100–200 MHz [1], and for NbN it is predicted to be of several GHz [2]. With higher T_c, a higher operating temperature can be used, yielding shorter τ_{e-ph}. NbN mixers have so far demonstrated 1 GHz bandwidth [16], [17].

The bolometer resistance may be adjusted by adding parallel strips and by changing the strip length. For $f_{\rm LO}, f_s > 2\Delta/h$ the absorption of microwave power by the normal electrons in the HEB is essentially frequency independent. Therefore, these mixers will operate up to RF frequencies well into the THz range, i.e., much higher than the gap frequency of the superconductor. Previous work has, in fact, shown mixing at 1.5 THz [1].

III. Mixer Theory

Fig. 2 shows an equivalent circuit of the mixer, where the device is biased by a constant dc current. The heating of the electrons by RF or dc power will increase the device resistance. The signal, when beating with the LO, will cause a modulation of the device resistance at $f_{\rm IF}$. An IF voltage will appear across the device, causing a current through the IF load resistance. This current will add to the dc bias current through the mixer device, and will create feedback. Note that the resistance of the device is time dependent only for the IF frequency. At the RF frequency the device impedance is assumed constant since these frequencies are higher than the inverse response time of the electron subsystem. If this statement were not true, an image current would be created, and a more complicated theory would be necessary. We have failed to detect any sidebands at $(f_{\rm LO} \pm f_{\rm IF})$ when a strong signal was applied in the IF-band. Our observation supports this feature of the device model.

Equivalent expressions for the conversion gain have been derived previously for three different types of HEB mixers: 1) the superconducting HEB by employing an energy balance equation [1]; 2) the hot electron InSb [8]; and 3) 2-DEG mixers [11], [12], the latter two directly from the circuit model of Fig. 2. A basic assumption is that the dc and RF power dependencies of the device resistance are equal (the time average RF power is employed since $f_{\rm RF} \gg (2\pi\tau_{e-ph})^{-1}$). Assume that in the bias point of the device we have $V = V_o$ and $I = I_o$. Defining the device dc-resistance as $R_o = V_o/I_o$ one obtains [8]

$$G = \frac{P_{\rm IF}}{P_s} = 2C_o^2 \frac{P_{\rm LO}P_{\rm dc}}{(R_L+R_o)^2}\frac{R_L}{R_o} \cdot \left(1 - C_o\frac{P_{\rm dc}}{R_o}\frac{R_L-R_o}{R_L+R_o}\right)^{-2} \qquad (1)$$

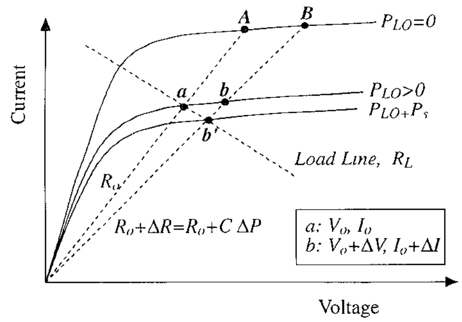


Fig. 3. Schematic illustration of the behavior of the I–V-characteristic and bias-point when LO- and RF-power is applied to the bolometric device.

where $C_o = dR_o/dP_{\rm device}, R_L$ is the IF load resistance, $P_{\rm IF}, P_s, P_{\rm LO}$ and $P_{\rm dc}$, are the IF-, signal-, LO-, and dc-power dissipated in the device, $P_{\rm device} = P_{\rm LO} + P_{\rm dc}$.

For the types of I–V-characteristics which apply to InSb and 2-DEG mixers, dV/dI is always positive, and an optimum conversion gain of −6 dB can be derived from (1), see [3], [8]. The I–V-characteristic of the superconducting HEB may have $dV/dI < 0$, and a re-interpretation of (1) leads to a prediction that $G > 0$ dB is possible, as shown in the next three sections. Conversion gain of >−6 dB for superconducting HEB's was originally suggested in [1].

IV. Relating the Conversion Gain to the I–V-Characteristic

A basic assumption in our model is that the low frequency resistance of the device is dependent on electron temperature. Hence, if the temperature increases linearly with dissipated power, we have $\Delta R = C_o \cdot \Delta P$. This assumption has certain interesting implications; consider the I–V-characteristic shown in Fig. 3. For a given device resistance $R_o = V_o/I_o$, the device temperature must be constant, i.e., the total power dissipated must be the same. When the device is pumped with a certain LO-power, the I–V-characteristic will change, as shown in Fig. 3. The above argument indicates that in points A and a we have the same total dissipated power; in A only dc power, and in a LO power plus dc power. If we increase either the LO power or the dc power, or both, the resistance will increase, and become $R_o + \Delta R$.

We may now calculate ΔR and ΔP (constant LO power)

$$\Delta R = \frac{V_o + \Delta V}{I_o + \Delta I} - \frac{V_o}{I_o} \approx \frac{V_o}{I_o}\frac{\Delta I}{I_o}\left(\frac{\Delta V}{\Delta I}\frac{I_o}{V_o} - 1\right) \quad (2)$$

$$\Delta P = (V_o + \Delta V)(I_o + \Delta I) - V_o I_o \approx V_o \Delta I \cdot \left(\frac{\Delta V}{\Delta I}\frac{I_o}{V_o} + 1\right). \quad (3)$$

From (2) and (3) we get

$$C_o = \frac{dR}{dP} = \frac{1}{I_o^2}\frac{\frac{\Delta V}{\Delta I}\frac{I_o}{V_o} - 1}{\frac{\Delta V}{\Delta I}\frac{I_o}{V_o} + 1} = \frac{R_o}{P_{\rm dc}}\frac{\left(\frac{dV}{dI}\right)_{\rm dc} - R_o}{\left(\frac{dV}{dI}\right)_{\rm dc} + R_o} \quad (4)$$

where $(dV/dI)_{\rm dc}$ is the differential resistance of the pumped I–V in the bias point determined by dc bias and LO power. Since it makes no difference whether ΔP is dc power or a combination of dc and LO power, we may use this value for C_o in (1), to obtain

$$G = \frac{1}{2}\frac{P_{\rm LO}}{P_{\rm dc}}\frac{R_L}{R_o}\left(1 - \frac{R_o}{(dV/dI)_{\rm dc}}\right)^2\left(1 + \frac{R_L}{(dV/dI)_{\rm dc}}\right)^{-2}. \quad (5)$$

There is still another relationship between the $P_{\rm LO}, P_{\rm dc}$ and I_o that can be used to rewrite the expression for conversion gain. For constant $R_o = V_o/I_o$, the dissipated power in the device must be constant. This means that if we have a bias current I_{oo} for the *unpumped* I–V, and I_o for the pumped I–V, and we have $P_{\rm LO} + R_o I_o^2 = R_o I_{oo}^2$. Then (5) can be rewritten as

$$G = 2\left(1 - \frac{I_o^2}{I_{oo}^2}\right)\frac{(C_o I_o^2)^2}{(1 - C_o I_o^2)^2}\frac{R_o \cdot R_L}{\left(R_L + \left(\frac{dV}{dI}\right)_{\rm dc}\right)^2} \quad (6)$$

from which it is seen that $R_L = (dV/dI)_{\rm dc}$ in the bias-point with LO yields maximum gain.

These equations for the conversion gain are valid for low intermediate frequencies, where the finite relaxation time of the device has no influence on the conversion gain. Notice that the proposed theoretical model does not require C_o to be independent of R_o. If C_o is independent of R_o and bias, the I–V satisfies the classical bolometer equation, $V_o/I_o =$ const. $+ C_o P$, where P is the dissipated power in the bolometer [18]. See Appendix for further details.

V. Predicting the Conversion Gain

The differential resistance can be predicted if $C_o I_{oo}^2$ is known. From (4) the differential resistance can be expressed as

$$\frac{dV}{dI} = R_o \cdot \frac{1 + C_o I_{oo}^2}{1 - C_o I_{oo}^2}. \quad (7)$$

Hence a negative resistance corresponds to $C_o \cdot I_{oo}^2 > 1$. Using (6) we have calculated the conversion gain vs. $(I_o/I_{oo})^2 = P_{\rm dc}/(P_{\rm dc} + P_{\rm LO})$ (see Figs. 4 and 5).

The fundamental limit of −6 dB gain for InSb mixers [8] is obviously not valid for the superconducting HEB mixers since it is possible to achieve positive conversion gain with negative differential resistance of the unpumped I–V-curve. From Figs. 4 and 5 it is seen that larger conversion gain is available for $R_L/R_0 > 1$ than for $R_L/R_o < 1$.

As mentioned above the load resistance for maximum gain is equal to the differential resistance of the I–V-curve in the bias point of the pumped mixer. It is also possible to find the optimum LO power (or $(I_o/I_{oo})^2$) for a given R_L/R_o or for R_L equal to the differential resistance of the I–V-curve in the bias point.

VI. Conversion Gain and IF Impedance Versus Intermediate Frequency

The coefficient $C_o = dR/dP$ can be expressed as $C_o = [dR/d\theta][d\theta/dP]$. Since the factor $d\theta/dP$ has a frequency dependence determined by the factor $1/(1 + j\omega\tau_\theta)$, and

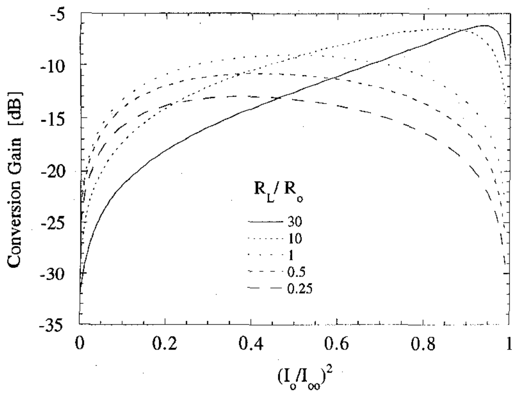


Fig. 4. Conversion gain versus $P_{dc}/(P_{dc}+P_{LO})$ for different IF load resistances. The differential resistance for the unpumped curve: $dV/dI=\infty$, or $C_oI_{oo}^2=1$ ($P_{dc}=R_oI_o^2$).

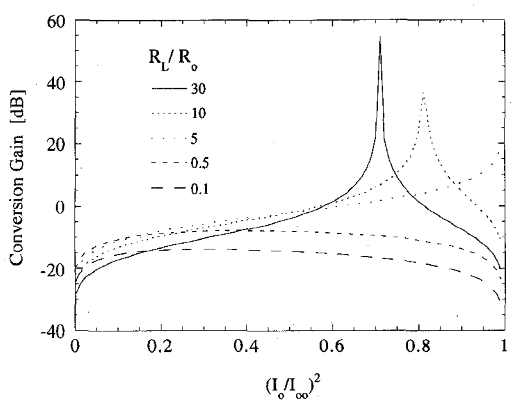


Fig. 5. Conversion gain versus $P_{dc}/(P_{dc}+P_{LO})$ for different IF load resistances. The differential resistance for the unpumped curve is negative: $dV/dI=-5\cdot R_o$, or $C_o\cdot I_{oo}^2=1.5$.

assuming that the term $dR/d\theta$ is simply a constant, we may conclude that

$$C_o(\omega)=\frac{Co(\omega=0)}{1+j\omega\tau_\theta}. \quad (8)$$

Using this expression for C_o in the equations for G above, the IF dependence of the conversion gain of the mixer can be expressed as

$$G=G(\omega=0)\cdot\frac{1}{1+(\omega\tau_{\text{mix}})^2} \quad (9)$$

where

$$\tau_{\text{mix}}=\frac{\tau_\theta}{1-C_oI_o^2\cdot\dfrac{R_L-R_o}{R_L+R_o}}. \quad (10)$$

We have introduced τ_{mix}, which determines the -3 dB bandwidth. It is possible to make the bandwidth wider by making the factor $C_oI_o^2\cdot(R_o-R_L)/(R_o+R_L)$ large ($R_o>R_L$). Furthermore, a value of $C_oI_o^2\cdot(R_o-R_L)/(R_o+R_L)$ close to 1 will lead to large gain (1), as well as a narrowing of the bandwidth. In practice, one would find a compromise between IF bandwidth and conversion gain. We can also conclude that it is undesirable to attempt to utilize a large conversion gain, a feature which the HEB mixer shares with other negative resistance devices.

The frequency dependence of the IF impedance can be derived in the same manner by using (8) and (4). We get

$$Z(\omega)=R_o\left(1+\frac{2C_oI_o^2}{1-C_oI_o^2}\frac{1}{1+j\omega\tau_{\text{imp}}}\right) \quad (11)$$

where

$$\tau_{\text{imp}}=\frac{\tau_\theta}{1-C_oI_o^2}. \quad (12)$$

In interpreting this equation we note that the finite response time of the heated electron medium, which gives rise to the bandwidth of the mixer (10), also makes the impedance at the intermediate frequency dependent at low frequencies. At the lowest frequencies, we expect the impedance to be given by (dV/dI) for the particular point on the I–V-curve, which should be real, while for frequencies much higher than $1/(2\pi\tau_{\text{imp}})$ we should expect a resistive impedance R_o as well. For intermediate frequencies, we know by the Kramers–Kronig theorem, that a reactive part of the impedance will develop, whenever the real part is frequency-dependent. These facts are consistent with (12). Note that τ_{imp} depends on the operating point, through I_o, in a slightly different manner than τ_{mix}. These phenomena were first demonstrated in Nb films in [19].

It is possible to rewrite the equation for conversion gain expressing the frequency dependence in terms of the electron temperature relaxation time τ_θ. Using a complex load impedance viz. $R_L\to Z_L=R_L+jX_L$ we obtain (13), shown at the bottom of the page.

Hence it should be possible to improve the bandwidth somewhat by a proper design of the load circuit Z_L (see Fig. 6).

VII. Theory of Noise in HEB's

The main noise of concern for the superconducting HEB is that which is directly emitted by the device at the IF frequency. The noise theory does not need to take into account correlation of noise contributions at a number of frequencies, as is the case for the Schottky-barrier mixer, for example. The IF output noise is primarily of two types [1], [13]: 1) Nyquist noise at the electron temperature, and 2) temperature fluctuation noise. The latter contribution is not usually important for microwave

$$G=\frac{2\left(1-\dfrac{I_o^2}{I_{oo}^2}\right)\dfrac{(C_oI_o^2)^2}{(1-C_oI_o^2)^2}R_o\cdot R_L}{\left|R_L+(dV/dI)_{\text{pumped}}-\omega\tau_\theta\cdot\dfrac{X_L}{1-C_oI_o^2}+j\left[X_L+\omega\tau_\theta\cdot\dfrac{R_o+R_L}{1-C_oI_o^2}\right]\right|^2} \quad (13)$$

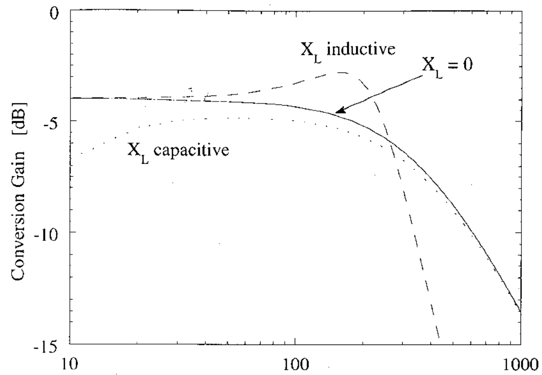


Fig. 6. Intrinsic conversion gain versus frequency for different load reactances. $\tau_\theta = 1.8$ ns.

or millimeter wave mixers, but is likely to be the dominant one in superconducting HEB mixers. Temperature fluctuation noise is well known from the theory of conventional (lattice) bolometers [20]. The electrons are assumed to form a subsystem in thermal equilibrium at the electron temperature, θ, while conducting heat to the lattice heat reservoir at temperature T. The corresponding thermal conductance is G_e [W/K]. It is known from thermodynamics that the average temperature of the electron subsystem will show spectral RMS fluctuations in a 1-Hz bandwidth given by [20]

$$\langle \Delta\theta^2 \rangle = \frac{4k_B\theta^2}{G_e}. \tag{14}$$

The thermal conductance is

$$G_e = \frac{C_e V}{\tau_\theta} \tag{15}$$

where V the volume of the superconducting strip. Since the resistance of the HEB depends on θ, and the device is biased with constant dc current, I_o, there will be a noise voltage developed across the device. We can derive the equivalent device output noise temperature, $T_{d,t}$, by equating the noise power due to temperature fluctuations to that from a Nyquist source at $T_{d,t}$

$$T_{d,t} = \frac{I_o^2 (dR/d\theta)^2 \theta^2}{R_o G_e}. \tag{16}$$

Here $(dR/d\theta)$ must be evaluated for actual mixer operating conditions. The total device output noise temperature will be

$$T_d \approx \theta + T_{d,t}. \tag{17}$$

This noise temperature can now be used in standard expressions for the mixer noise temperature, given the conversion loss, L. If typical values are inserted in (16), one finds that this term is likely to dominate. Note that $T_{d,t}$ is proportional to the electron temperature squared. Furthermore, θ is expected to be close to T_c. HEB's based on superconductors with higher T_c then should have higher noise. The θ-dependence may be compensated by the appearance of G_e in the denominator, however. As a rule, higher T_c materials have shorter relaxation times, resulting in larger G_e (see (15)). An optimum material in terms of the output noise temperature of the device can therefore not be identified. One must also consider the conversion loss which can be achieved, in order to minimize the receiver noise temperature. These matters are under investigation, but have not yet been resolved in detail. Another noteworthy consequence of the dominance of $T_{d,t}$ is that the output noise temperature is predicted to fall with IF frequency with a time constant similar to τ_{mix} (see (10)). If this can be verified, then one concludes that the bandwidth over which a given mixer noise temperature can be maintained is actually wider than $1/(2\pi\tau_{\text{mix}})$, since the device output noise decreases at the same time that the mixer conversion gain decreases. Measurements of the output noise versus IF-frequency have so far given inconclusive results.

In conclusion, one can roughly estimate a device output noise temperature for Nb in the range 25–100 K. This is consistent with measurements, as described in a later section.

VIII. Experiment

The HEB devices were made from 90–150-Å-thick niobium films, dc-magnetron sputtered on silicon substrates held at room temperature and patterned by conventional photolithography. The HEB's consisting of two parallel niobium strips, 1.5 μm wide and 7 μm long, had a normal resistance between 40 and 150 Ω, and were measured at 20 GHz signal and 1–1000 MHz intermediate frequency. T_c of the 90-Å-thick devices was about 4.3 K. This value is expected for thinner films [21]. However, our films are not passivated, and therefore include a 20–30-Å-thick oxide layer. The films also have a diffusion constant of about half the value of the constant given by [22].

The best conversion is achieved when the HEB mixer is biased in the lower part of the resistive region (Figs. 7 and 11). For no or small LO power, at low temperatures compared to T_c, we have observed negative differential resistance in the dc I–V-curve in this part of the resistive region (Fig. 11 unpumped curve at 2.1 K) similar to the behavior of self-heating phenomena, i.e., hot spots or resistive domains [23]. This indicates the possibility to obtain positive conversion gain.

Several samples (mainly 90 Å thick) have been measured with consistent results [3]. However, all figures in this section shows the measured data from one typical device. The results from mixing experiments are given in terms of the intrinsic conversion loss of the mixer. By intrinsic conversion loss we mean the ratio between the RF signal power from a monochromatic source coupled to the device to the IF signal power delivered to the IF load.

A. Determination of RF Coupling Loss

To determine how much of the applied RF power is coupled to the device it is assumed that the response of the device to RF power is equal to the response to dc power. Several I–V curves for different applied RF powers are recorded, with the RF power changed in steps of 0.5 dB. The coupling loss is then found as the ratio between the change in applied RF power and the corresponding change in dc power for the different

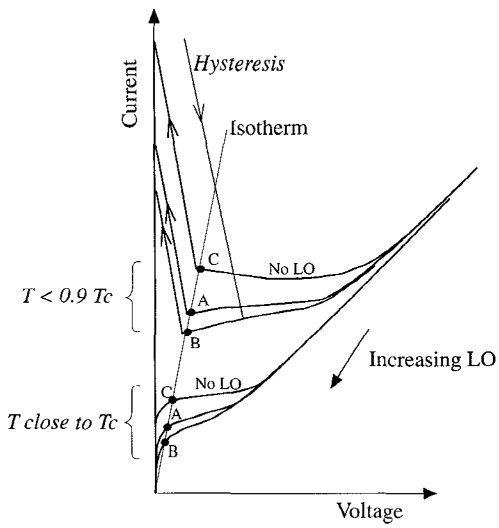


Fig. 7. Two sets of I–V curves for different temperatures and applied LO powers. For one set the physical temperature is close to T_c, and for the other set the temperature is less than about $0.9T_c$. For $T < 0.9T_c$ hysteresis can be observed.

curves. The points on the I–V curves used for deriving the dc powers have to be along the isotherm which passes through the bias point, typically A in Fig. 7.

We derive the RF Coupling Loss from

$$L_c = \frac{P_{\mathrm{LO},B} - P_{\mathrm{LO},A}}{P_{\mathrm{dc},A} - P_{\mathrm{dc},B}} \tag{18}$$

where $P_{\mathrm{LO},A}$ and $P_{\mathrm{LO},B}$ are determined at the coaxial connector input to the liquid helium dewar. The coupling loss thus includes the attenuation in the input coaxial cable, and the microstrip circuit to which the device is connected, as well as reflection loss of the device impedance with respect to the 50-Ω microstrip circuit. The attenuation excluding reflection loss at the device has been estimated to about 3 dB.

The measured reflection loss at the device is bias dependent as shown in Fig. 8. If we assume that the device impedance is R_o, then we can calculate the reflection loss as shown in the full-drawn curve in Fig. 8. At about 3-mV bias, R_o, as defined above, is close to 50 Ω, and the reflection loss very small. The measured coupling loss shows a similar trend, but is about 1.5 dB higher at low bias voltages. The device resistance at 20 GHz thus is fairly close to R_o. If the frequency of the RF radiation were higher than the gap frequency in the entire strip, then we would expect the resistance to be equal to R_N, and independent of the bias. This is likely to be the situation in THz versions of the HEB mixer, but is not the case in the 20-GHz prototype. The frequency and bias dependence of the device impedance thus requires further study, as we attempt to increase the frequency at which the HEB mixer operates.

B. Intrinsic Conversion Loss

The lowest intrinsic conversion loss of several investigated devices was around 1 dB at 20 MHz IF and operating temperatures between 4.2 and 2.1 K. The IF of 20 MHz was chosen for the experiments since it is well below the cut-off frequency for hot electron effects (compare Fig. 12).

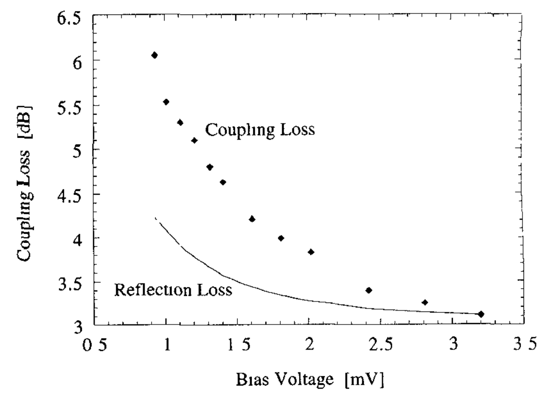


Fig. 8. Measured 20-GHz RF coupling loss (points) versus bias voltage, and calculated reflection loss under the assumption that the bolometer microwave resistance is equal to its dc resistance.

In Fig. 9 one can see how the conversion loss depends on bias voltage and temperature. The lowest conversion loss is found for the smallest bias, i.e., the lowest bolometer dc resistance. Besides dc-bias, $\Delta T_c, T_c$, physical temperature, LO power, and the number of strips have important effects on the mixer performance. So far we have not carried out a complete investigation of the variation of conversion loss with these parameters. However in general terms we have found the following:

1) The best conversion gain is found for samples with the narrowest superconducting-normal transition width ΔT_c, i.e., dR/dT is large.
2) The change in conversion gain upon a change in LO power by 1 dB is less than about 0.5 dB. See Fig. 10.
3) From measurements for a device temperature of $T = 2.1$ K, and $L \approx 1$ dB, we have estimated that the change in conversion loss is less than 1 dB for a change in temperature of about 0.15 K, or $dL/dT = 7$ dB/K. For $T = 4.2$ K $dL/dT = 50$ dB/K.
4) For any temperature in the interval 2.1 to 4.2 K the conversion loss will be the same after optimization of bias and LO power.
5) Adding more parallel strips results in a smoother I–V curve and a more stable output signal, hence a better agreement between measurement and theory.

In Fig. 9, the measured intrinsic conversion loss is compared with the theoretical conversion loss calculated from (1) and (6) for two temperatures, 4.2 K ($\approx$ 0.05 K below T_c of the thin strips) and 2.1 K. Our model agrees within a couple of dB with measurements. The I–V curves of the mixer at 4.2 and 2.1 K are shown in Fig. 11. $P_{\mathrm{LO}}, R_o, P_{\mathrm{dc}}, I_o, I_{oo}$, and $(dV/dI)_{\mathrm{dc}}$ were obtained from the dc I–V plot. Absorbed pump and signal powers, about −44 dBm and −72 dBm respectively, are typical values in our experiments. It is interesting that the conversion loss at 4.2 and 2.1 K is roughly the same if R_o is the same.

Note in Fig. 11 that for low bias and no LO, a region with negative resistance is clearly observed. A probable explanation for the occurrence of the negative resistance is the following. When the bias is lowered, the heating is lowered and the

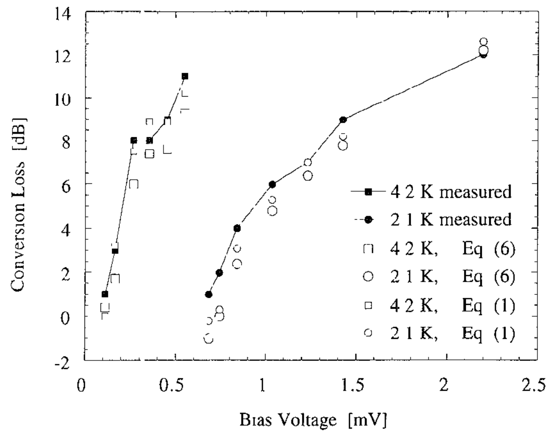


Fig. 9. Experimental and calculated intrinsic conversion loss versus bias voltage and temperature. The LO power is optimized at the lowest bias voltage and kept constant when bias is changed. In the modified Arams expression (6), the slope of the unpumped curve is calculated from the slope of the pumped curve.

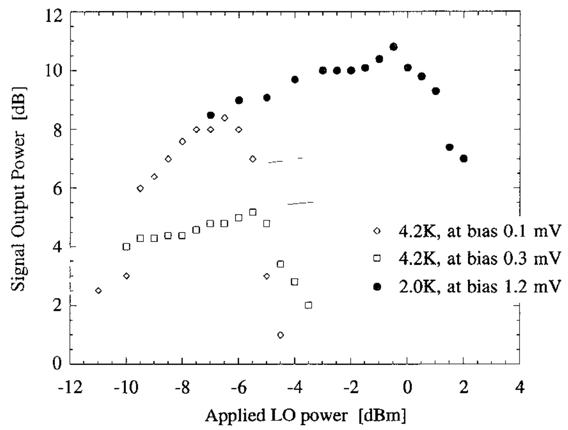


Fig. 10. Output power versus applied LO power.

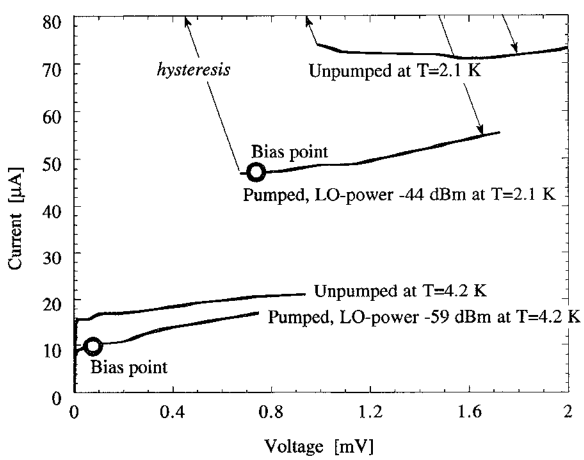


Fig. 11. Measured I–V curves with and without pump power at 4.2 and 2.1 K.

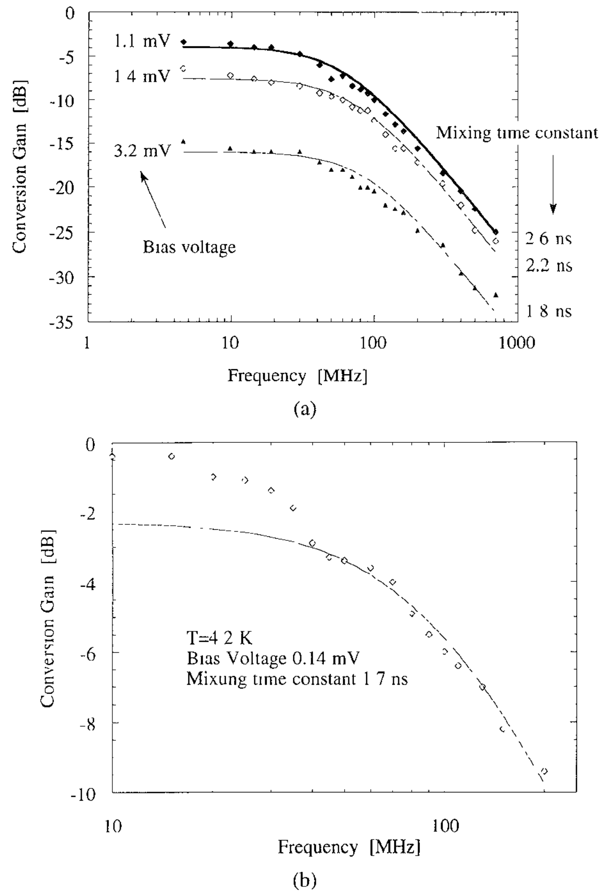


Fig. 12. (a) Conversion gain versus frequency and bias, showing an increased mixing time constant for lower bias at 2.1 K, and (b) a distinct low-frequency response below 40 MHz at 4.2 K.

temperature of the Nb lattice is lowered as well. This should lead to an increase in the maximum current density, and we consequently will see a negative differential resistance. The time constant for the combined heating phenomenon will be determined by τ_{e-ph}, since $\tau_{ph-s} \ll \tau_{ph-e}$.

C. IF Bandwidth

The conversion gain has dropped 3 dB at around 80 MHz IF, see Fig. 12(a) for $T = 2.1$ K. For larger bias voltages than for maximum conversion, there is a small increase in bandwidth, i.e., the mixer time constant will be slightly shorter with larger bias voltage. Assuming the relaxation time constant $\tau_\theta = 1.78$ ns, and calculating $C_o I_o^2$ and R_o from the dc I–V-characteristic, we obtain theoretically from (10) $\tau_{\rm mix} =$ 2.6, 2.14, and 1.83 ns, respectively, for the three bias voltages 1.1, 1.4, and 3.2 mV, respectively, in excellent agreement with the experimental time constants. In Fig. 12(b) it can be seen that at 4.2 K there is a response at lower IF around 1–40 MHz which is difficult to distinguish at 2.1 K. The low frequency response is due to influence from slower processes, such as the relaxation of the normal domain size.

It should be possible to achieve a more uniform absorption of RF power in the strips by: 1) Application of a magnetic field, which reduces the energy-gap of the superconductor below $f_{\rm RF} \cdot h$, allowing absorption also in the superconducting regions, 2) Use of higher LO and signal frequencies, comparable to or above the gap frequency of the superconductor at the actual temperature, and 3) reduction of the energy-gap by increasing the lattice temperature to a value closer to T_c. It is noteworthy that a higher-frequency HEB mixer is expected to have a more ideal performance than the low-frequency prototype we have investigated.

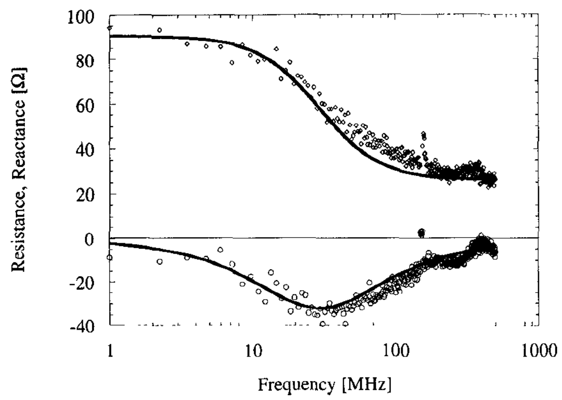


Fig. 13. The measured bolometer impedance for (dots). The solid lines show the real and imaginary part of the circuit in Fig. 14. R_o is 26 Ω and $(dV/dI)_{dc} = hbox108\,\Omega$. Optimum LO power is applied. $T = 2$ K.

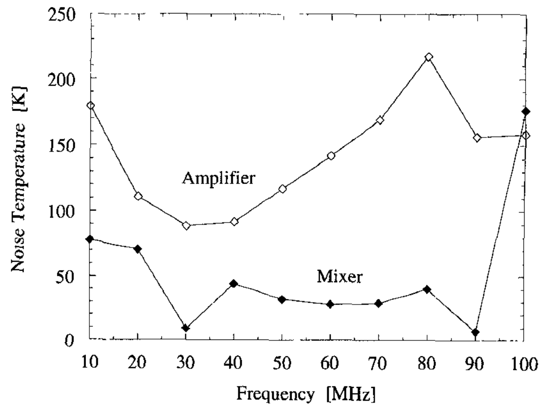
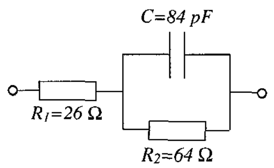


Fig. 15. Output noise temperature of mixer, and amplifier noise when connected to the mixer. $R_o = 26$ Ω, $T = 2.1$ K. P_{LO} is optimized for maximum conversion gain.



$$Z = R_l + \frac{R_2}{1+j\omega R_2 C}$$

Fig. 14. Equivalent circuit for the bolometer. The values for R_1, R_2, and C apply to the sample and were derived from the measured bolometer impedance data in Fig. 13.

D. IF Impedance

The IF-impedance of the device has been measured between 1 and 500 MHz (see Fig. 13). There is a transition of the impedance in the range 10 to 100 MHz from 90 Ω, which is close to the dc differential resistance (108 Ω) in the bias point, to 26 Ω, which is the dc resistance in the bias point. The impedance could not be measured for lower biases, i.e., not at the point for optimum conversion, due to noise from the network analyzer.

Fig. 14 shows the equivalent circuit of the device, which satisfies (11), as well as the measured data in Fig. 13. The electrical capacitance, C, is used to simulate the thermal inertia of the device due to the finite electron temperature relaxation time. The time constant $\tau = R_2C$ is 5.37 ns. Using the value for $C_oI_o^2$ in the bias point and (12) we find a value for $\tau_\theta =$ 1.89 ns, to be compared to the number obtained from the conversion loss measurement ($\tau_\theta = 1.78$ ns). In fact using a slightly shorter τ would yield an even better agreement with measurements (compare Fig. 13).

E. Noise

The receiver noise was measured with a solid state noise source (3815 K) coupled to the device together with the LO signal using a directional coupler [24]. The IF signal was amplified using a low-noise amplifier with a noise temperature T_A according to Fig. 15. The IF power was registered using a spectrum analyzer. The influence on the system noise of the components between the noise source and the mixer had to be subtracted, introducing a considerable uncertainty in the determination of the mixer noise temperature. The observed total receiver noise temperature was between 470 and 690 K. A DSB *mixer noise* temperature between 80 and 450 K can be derived from these values, given that the intrinsic conversion loss in the bias point is 7 ± 1 dB.

The *output noise* of the mixer was measured by comparing the output power from the pumped device with a hot and cold load (295 and 77 K, respectively). The output signal was amplified by an LNA (with noise temperature T_A in Fig. 15) and fed through a receiver (bandwidth 10 MHz) and registered on a power meter. Since no isolator was available for isolation of the IF amplifier, a coolable load was fabricated from lumped resistors and a capacitor (see Fig. 14) to simulate the device impedance. The impedance of this load was checked with a network analyzer, and agreed well with the measured data for the device over the frequency range 1–100 MHz (see Fig. 13). The output noise from the mixer at intermediate frequencies of 20–90 MHz was found to be 30–50 K (see Fig. 15). For this measurement, it was not possible to bias the device for optimum gain. However, the optimum LO power was supplied. In this bias-point, the intrinsic conversion loss was 7 dB, and we can make an estimation of the DSB mixer noise temperature to be between 80 and 150 K, based on the measured output noise temperature. The estimated DSB mixer noise temperature is consistent with the above measurements of the mixer receiver noise.

IX. Conclusion

We have presented a simple phenomenological theoretical model for the bolometric mixer. This model allows us to derive conversion loss and bandwidths from knowledge of the unpumped and/or pumped I–V characteristic. We do see a very low conversion loss, which, if below −6 dB, is related to negative resistance in the unpumped I–V. Our measurements agree well with the theoretical model. It is predicted that conversion gain can be achieved, if a negative differential resistance is available for the unpumped I–V and if a proper IF impedance load is chosen. The penalty for larger gain is smaller IF bandwidth.

The measured noise is quite low. We see noise temperatures out from the device typically around 50 K, and a double

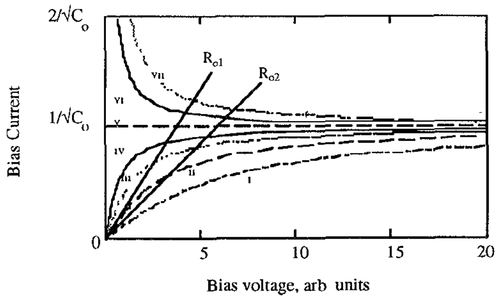


Fig. 16. I–V curves calculated from (A4). The curves i–vii represent decreasing values for $R_{oo}+C_oP_{\rm LO}$. For curve v, $R_{oo}+C_oP_{\rm LO}=0$, for curve vi and vii, $R_{oo}+C_oP_{\rm LO}<0$. It is seen that e.g., along the line representing R_{01}, the differential resistance gradually changes from a negative value to positive upon increasing LO power.

sideband noise temperature expected for an optimized mixer is calculated to about 100–150 K.

The impedance of the device at 20 GHz is bias-dependent, which suggests that this frequency is lower than the frequency corresponding to the energy gap of the superconducting regions of the device. The frequency dependence of the IF-impedance and conversion gain has been modeled and measured. The measurements are consistent with an electron energy relaxation time of 1.8–1.9 ns. This means that the maximum IF frequency for this particular device is about 100 MHz. It is pointed out that the IF-frequency dependence of the conversion gain can be improved by a proper IF load circuit.

Acknowledgment

The authors thank Prof. E. Gershenzon and Prof. G. Gol'tsman for numerous fruitful discussions.

Appendix
The Classical Bolometer I–V-Characteristic

The differential resistance can be predicted if C_oI^2 is known. From (4) we get

$$\frac{dV}{dI}=\frac{V}{I}\cdot\frac{1+C_oI^2}{1-C_oI^2}. \tag{A1}$$

The theoretical model presented *does not require the same* C_0 *for all different values of the bolometer resistance* $R_o=V/I$. However, if we assume that C_0 *indeed is the same for all* R_o we can integrate (A1). We obtain

$$\frac{V}{I}=\frac{R_{oo}}{1-C_oI^2} \tag{A2}$$

where $R_{oo}=dV/dI$ for $V=I=0$. (A2) can be rewritten as

$$\frac{V}{I}=R_{oo}+C_o\cdot P_{\rm dc} \tag{A3}$$

which is identical to the classical bolometric formula ($P_{\rm dc}=IV$).

If part of the dc-power, ($P_{\rm dc}$) in (A3), is replaced by LO-power, we have

$$\frac{V}{I}=\frac{R_{oo}+C_oP_{\rm LO}}{1-C_oI^2}. \tag{A4}$$

Fig. 16 shows I–V curves obtained using (A2) for different $R_{oo}+C_oP_{\rm LO}$. For $C_oI^2\rightarrow 1, V\rightarrow\infty$. The curves with positive differential resistance requires $C_oI^2<1$, and $R_{oo}>0$, while those with negative differential resistance requires $R_{oo}<0$ and $C_oI^2>1$.

Obviously the model also in this case allows for negative differential resistance, a feature which is necessary condition to achieve a conversion loss less than 6 dB, or even conversion gain.

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